STUDY REPORT FOR THE DEVELOPMENT OF TECHNIQUES TO AUTOMATICALLY ACQUIRE THE CARRIER OF AM OR PM SIGNALS

15 JUNE 1967

CONTRACT NO. NAS5-10075

PREPARED BY

ELECTRAC, INC.

1614 ORANGETHORPE WAY

ANAHEIM, CALIFORNIA

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GREENBELT, MARYLAND

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SUMMARY

This report first analyzes some of the problems of fully automatic acquisition of the true carrier frequency of AM and PM signals by a phase-lock loop system operating down to its theoretical threshold. It has been found that for AM, the carrier can be determined from the knowledge that the spectrum is symmetrical. For the PM case, however, no theory could be advanced which assures a basis for completely satisfying the originally sought goals. A simplified problem is secondarily considered which calls for semi-automatic acquisition of the carrier. Here, a system is recommended which will permit the operator to more quickly acquire the carrier and to thereafter reduce or eliminate the need for further monitoring time on his part.

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I -- INTRODUCTION

This study was undertaken with the purpose of determining the best possible hardware approach to the GSFC acquisition problem as outlined in the Specification S-523-P-5, and as described in Electrac Proposal E.I. 391 dated 2 June 1967.

The basic problem may be summarized as being the acquisition of the true carrier frequency of amplitude or phase-modulated (AM or PM) signals by the phase-lock loop systems now employed in the GSFC tracking stations.

The report first analyzes the general problem presented by the specifications and finds that a complete answer cannot be supported theoretically.

Secondarily, a system is recommended which solves a less severe problem -- that of speeding up the initial acquisition process (even if not to the correct true carrier) and then provides a monitoring and warning system if more than one spectral line exists in the band of interest. This system, while far from solving the general problem posed here, may have practical benefits to the station operator in reducing and making his time more effective.

II -- DISCUSSION

2.1 Phase-Lock Loop Problems

The phase-lock loop used to coherently demodulate the PM (or AM) signals can be locked to any spectral line contained in the complex radio frequency spectrum. This spectral line may be the carrier frequency or one of any number of sidebands resulting from periodic frequency components (spectral lines) of the complex modulating signal. These lines, in general, vary slowly with time, and the phase-lock loop system is capable of tracking and measuring their frequency and amplitude even when the signal-to-noise ratio of the complex signal may be considerably less than unity. However, the system does not have any way of distinguishing which one of these is the true carrier and which is a sideband. Lock to a sideband renders the resulting demodulated signal meaningless and, therefore, acquisition of the correct carrier frequency is essential.

2.2 Classification of Information

It is evident that a priori information is required to determine the frequency of the carrier. For the purpose of this report, we will arbitrarily classify two types of available knowledge which may or may not be available at the time of acquisition. These are:

1. General knowledge of the modulation technique (i.e. PM, SSBSC, AM, etc.).

2. Detailed knowledge of the modulating signal structure (modulating frequency, spectrum, modulation indices, etc.).

The specifications list the following information:

A. "AM and PM only - no FM" (This is interpreted as meaning no dc FM or that the phase modulating signal has no component of the form kt.)

B. "At least 10 percent of the received power in carrier"

C. "No sideband component within three tracking bandwidths of the carrier"

D. "Full RF spectrum contained within pre-detection bandwidth"

E. "Total Doppler less than 200 kHz and maximum Doppler rate less than 2.5 kHz/sec."

Statements A and possibly B would represent Type 1 information; statements C and D applying to Type 2 information, and statement E is not applicable.

We now consider several types of modulation techniques to determine what knowledge is necessary for proper acquisition. Although not required, it is useful to first consider a suppressed carrier with a pure sinusoidal modulating signal. Here, the RF spectrum consists of a single spectral line not at the carrier frequency. In this example, both Type 1 and Type 2 information are required -- without <u>both</u> it is impossible to determine the carrier.

Next, consider AM, which is a requirement in this study. Because the sidebands are always symmetrically displaced about the carrier, the carrier can always be determined from its spectrum. Only Type 1 information has been required, and a device which can perform a spectral analysis and then determine the point of symmetry is all that is needed.

Also of interest, but not required here, is Suppressed Carrier AM. Here, we can determine the carrier from the sidebands alone. It is always possible to represent the signal by

$$e\ (t) = a\ (t)\ \mathrm{Sin}\ W_o t$$

If we square the signal and observe the second harmonic component,

$$e^2 = a^2 (t) \sin^2 W_o t$$

$$= \frac{a^2 (t)}{2} (1 + \sin 2W_o t)$$

Thus, a discrete line of amplitude a^2 occurs at the second harmonic of the carrier. Further, the fundamental and second harmonic bands have symmetrical spectra about the carrier. Thus, with Type 1 information only, we could determine the carrier frequency by the squaring process and analysis of the spectrum to determine the center of symmetry.

We will now consider the PM or FM only (no AM) case. If only Type 1 information is available, the general solution is theoretically possible by use of a frequency discriminator which can evaluate the mean value of frequencies. This follows because of the mentioned interpretation of "no FM" being present. Without this interpretation, the determination of the carrier is impossible without the addition of Type 2 information.

Of particular interest to us, for the PM or FM case, is the proposed use of the spectrum of the RF signal to determine the true carrier frequency. It is well known that an infinite number of spectral lines are generated in FM or

PM, and that for the simple special case of a single sinusoid modulating signal (Type 2 information), the carrier will occur at the point of symmetry. However, two sinusoids which are harmonically related can clearly provide an unsymmetrical spectrum. This is illustrated in Figure 1. Note in Figure 1a and 1c the considerable amount of non-symmetry. Further, calculation of the center of power of the spectrum for these examples show it to be considerably displaced from the true carrier.

It is concluded in Reference 1 that sideband symmetry about the carrier occurs only when the polarities of the modulating wave have symmetrical waveshapes or that symmetrical modulating signals produce symmetrical frequency modulated signal spectra and unsymmetrical modulating signals produce unsymmetrical spectra. Also, it is concluded that the energy in the sidebands tends to distribute itself in accordance with the shape of the modulating wave. One is forced to the final conclusion that spectral analysis does not offer a general solution.

2.3 Frequency Discriminator

As has been brought out, the use of a wide band frequency discriminator capable of giving the mean value of the input frequency appears to offer a theoretical means of establishing the true carrier frequency. The obvious

1. Giacoletto, "Generalized Theory of Multitone Amplitude and Frequency Modulation", IRE Proceedings, (July, 1947), V. 35, p. 680.

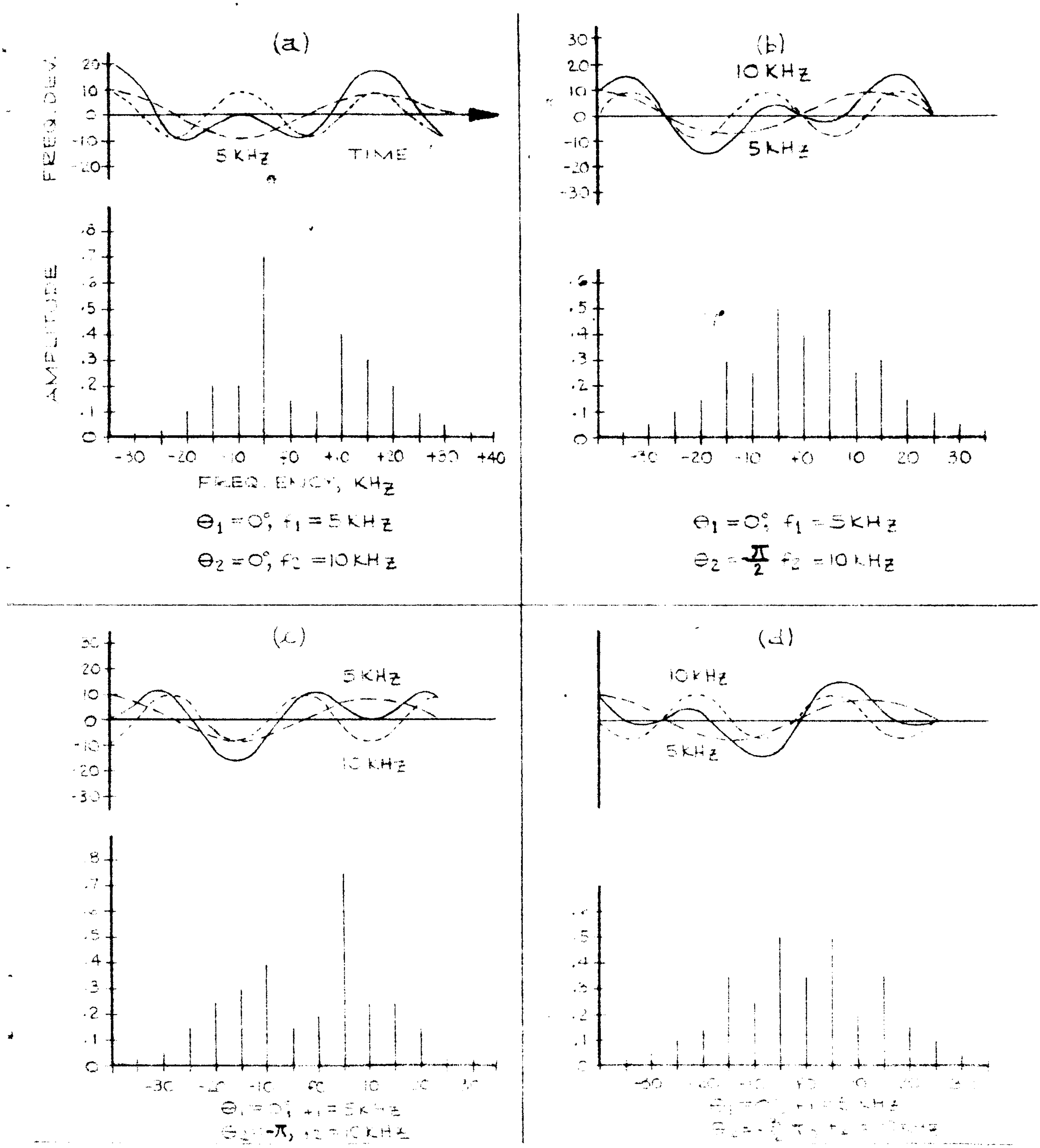

(a)
FREQ. DEV.
5 KHz
TIME
AMPLITUDE
FREQUENCY, KHz
$\Theta_1 = 0°,\ f_1 = 5$ KHz
$\Theta_2 = 0°,\ f_2 = 10$ KHz
(b)
10 KHz
5 KHz
$\Theta_1 = 0°,\ f_1 = 5$ KHz
$\Theta_2 = \frac{\pi}{2}\ f_2 = 10$ KHz
(c)
5 KHz
10 KHz
$\Theta_2 = -\pi,\ f_2 = 10$ KHz
(d)
10 KHz
5 KHz

FIGURE 1

mechanism for this device is a conventional automatic frequency control (afc) loop. This idea has been utilized in the current Diversity Locked Phase Demodulator (Contract NAS5-9864) as an anti-sideband lock scheme. A description of this system and the design evaluation report for it are included in the Appendix.

It was recognized in the referenced report that the system would be limited to ODB carrier-to-noise at the crystal discriminator used in the Model 315A. Considerable work was subsequently expended to obtain the necessary symmetry in that unit. Open to question here is whether or not a significantly better discriminator could be built and utilized in this same basic system to give higher performance. The possible merits of this approach will now be considered.

Theory indicates two effects occur as the carrier-to-noise power at the discriminator is reduced: from positive to negative carrier-to-noise ratio for the case of a wide input bandwidth (200 kHz, full spectrum width of received signal) and a relatively narrow post detection bandwidth:

1. The discriminator noise spectrum increases to a constant value of noise power density, and changes from a triangle shaped spectrum to a flat noise spectrum.

2. The output signal power, the dc component of which we require here, is suppressed by a factor of the noise-to-carrier power ratio.

The noise power output from the discriminator, based on the axis crossing discriminator of Stumpers, Reference 2, is shown in Figure 2. The effect of noise in the afc system is to produce a fluctuation of the VCO about the correct value of the carrier frequency. Suppose, for example, that we require that the rms frequency accuracy of the VCO be equal to the receiver loop bandwidth to reduce the probability of locking to "sidebands components ... three times the particular tracking bandwidth being used". From Figure 2 we note:

$$\frac{N_o^2}{B_o \frac{(B_i)}{2}} = 1 \text{ for c/n small}$$

and letting

$$N_o = B_{TF} = 10 \text{ Hz (tracking filter bandwidth)}$$

$$B_o = \frac{N_o^2}{B_i/2} = \frac{(10)^2}{10^5/2} = .002 \text{ Hz}$$

2. Stumpers, F. L. H. M., "Theory of Frequency-Modulation Noise", IRE Proceedings (September 1948) pp 1081-1092.

DISCRIMINATOR OUTPUT NOISE FOR LOW CARRIER TO NOISE RATIOS & RELATIVELY NARROW OUTPUT BANDWIDTH

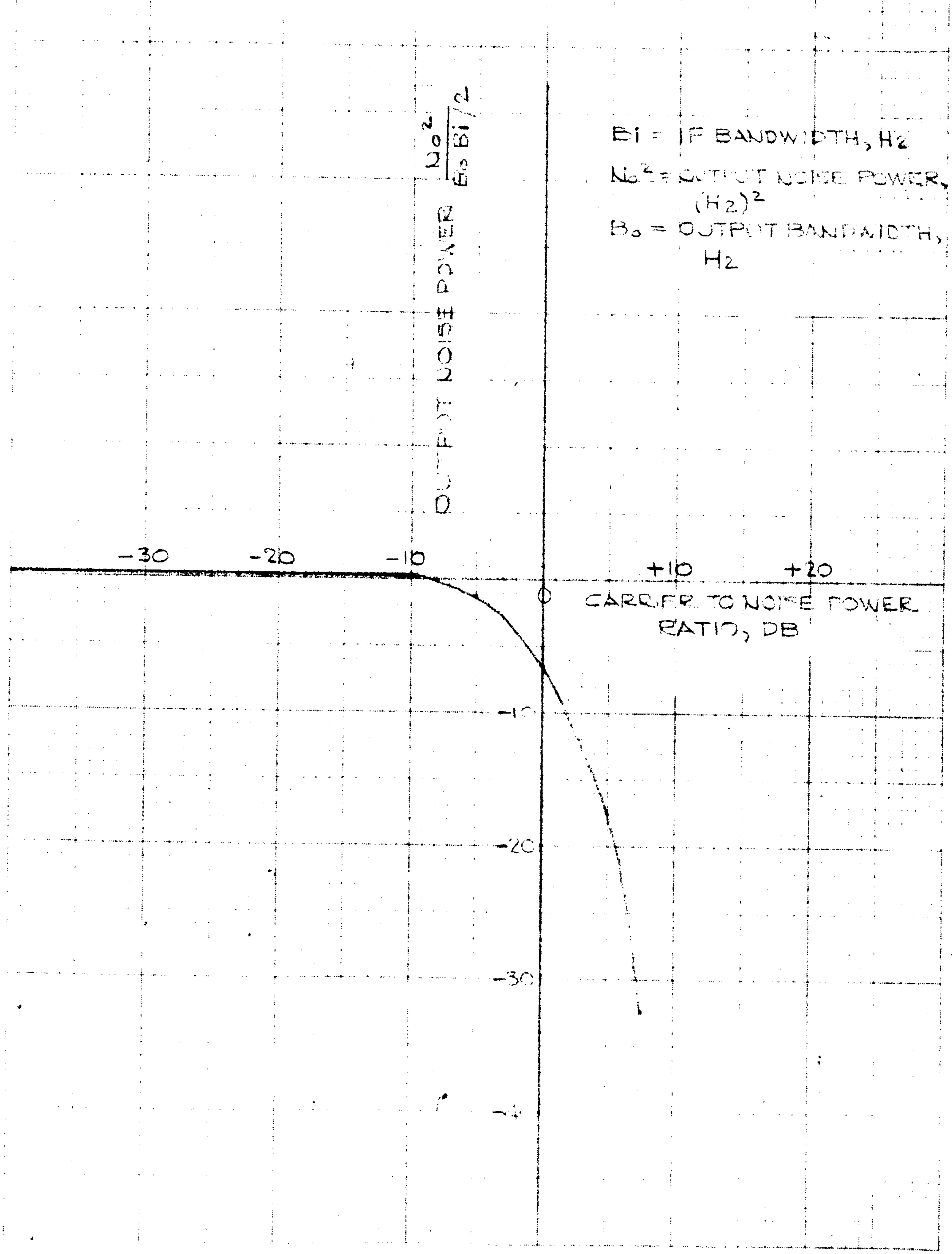


Figure 2

Values for other loop bandwidths for the afc systems are shown below:

Loop BW	B_o Discriminator Output Filter	Settling Time of afc loop
10	.002 Hz	500 seconds
30	.02 Hz	50 seconds
100	.2 Hz	5 seconds
300	2.0 Hz	0.5 seconds

Also, to get an idea of how fast the system would operate the settling time, approximated by $1/B_o$, has been tabulated.

From this, it is clear that for an ideal discriminator, a loop bandwidth of less than 30 Hz would have been impractical; and only the 300 Hz loop bandwidth appears capable of meeting the specified 1 second acquisition time (Paragraph 3.3.2 of the specifications).

Perhaps the more significant problem here is the signal suppression effect. The theoretical suppression effect is illustrated in Figure 3. Experimental values, taken on a production Electrac Model 315A Diversity Locked Demodulator, are also shown. The experimental values were taken by noting the voltage difference between the quartz-crystal discriminators dc output for two different known values of input frequency. The actual output voltage in practice changes with noise level due to the impossibility of achieving a perfect match in circuit diodes. At least from this

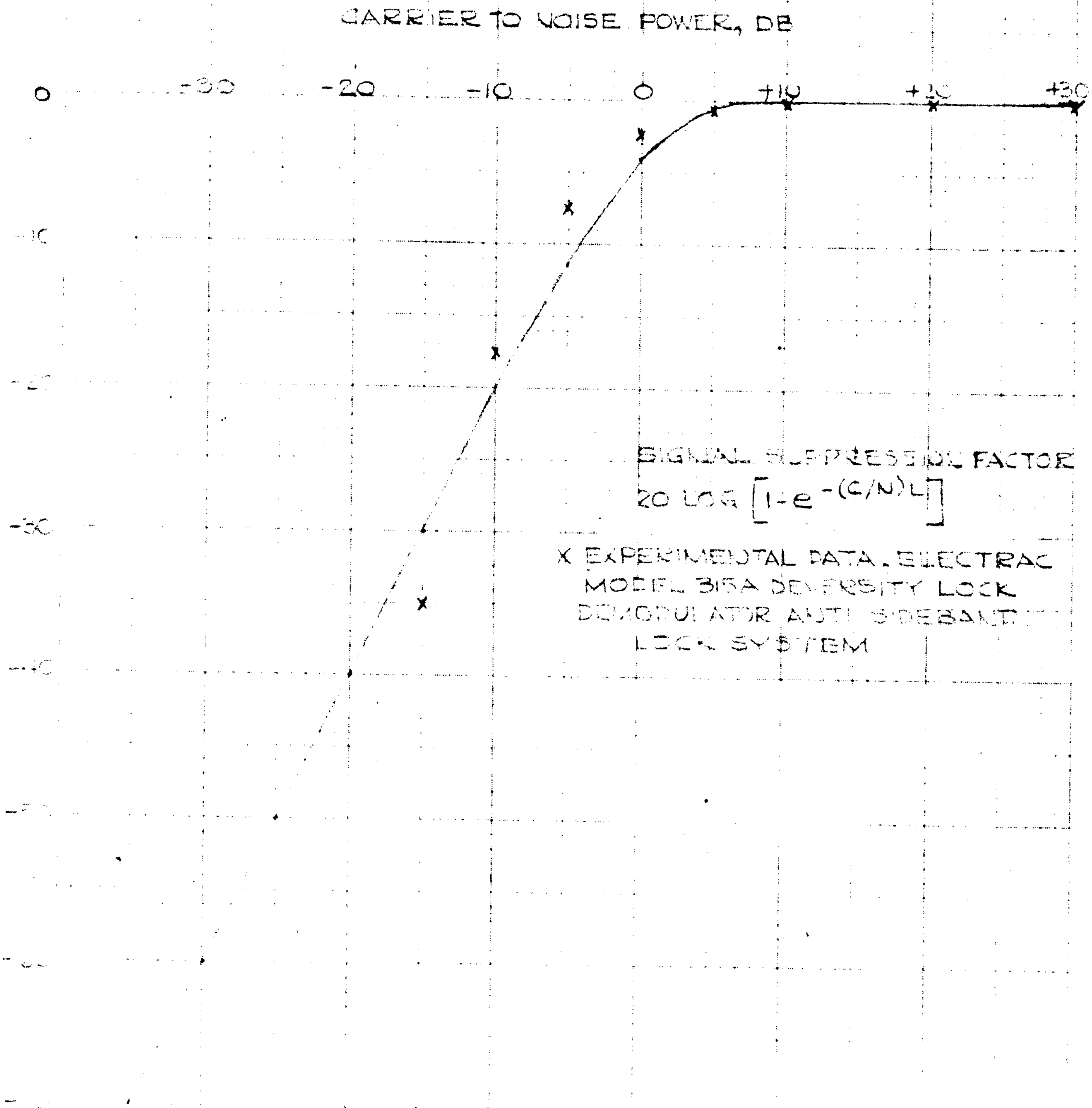

CARRIER TO NOISE POWER, DB
0
-30
-20
-10
0
+10
+20
+30
SIGNAL SUPPRESSION FACTOR
20 LOG [1-e -(C/N)L]
X EXPERIMENTAL DATA. ELECTRAC
MODEL 315A DIVERSITY LOCK
DEMODULATOR ANTI SIDEBAND
LOCK SYSTEM

Figure 3

experimental data, it does not appear practical to realize the discriminator below -10 db input carrier-to-noise ratios. In this respect, it appears that there could be value in expending effort towards the goal of achieving a well balanced discriminator. A digital type or an improved zero-crossing type might be employed. However, the possibility of achieving operation at the phase lock threshold looks rather remote.

The difficult in this approach is best brought out by example. Consider the problem implied in the 200 kHz discriminator requirement. For high signal-to-noise ratios, we would desire a linearity and long term zero frequency stability corresponding to approximately

$$\frac{10\text{ Hz}}{200{,}000\text{ Hz}} = 5 \times 10^{-5} \text{ or } .005\% \text{ of full scale output}$$

This is based on the assumption that all of the frequency spectrum is contained in the 200 kHz IF bandwidth, and that the discriminator must be sufficiently linear and stable to permit afc action to an accuracy of one times the loop bandwidth. We now, however, consider operation at negative carrier-to-noise ratios. Defining the phase lock threshold as 9 db in the loop bandwidth (20° rms phase jitter), we calculate the carrier-to-noise at the discriminator as +9 db $-10 \log (B_i/B)$. This is tabulated below.

Loop BW	SNR Loop BW	Carrier-to-Noise Ratio at Discr. Input	Signal Suppression	Factor
10 Hz	+9 db	-34 db	-34 db	50
30 Hz	+9 db	-29 db	-29 db	28
100 Hz	+9 db	-24 db	-24 db	16
300 Hz	+9 db	-19 db	-19 db	9

Note that for the worst case of 10 Hz, the sensitivity of the afc loop will be reduced to only two percent of its high signal level sensitivity. As a consequence, we must increase the balance and stability requirement by fifty times over the previously stated values, a feat not at all attainable by state of the art circuitry.

2.4 Multiple Discriminator System

A system wherein a series of discriminators having descending values of bandwidths are successively employed to "zero in" on the carrier has been considered. This system has been used to a limited extent (two bandwidths) on the Electrac Model 315A. The idea here is to reduce the impossible discriminator stability requirements as well as the signal suppression (increased carrier-to-noise ratio). This system has not been analyzed, to our knowledge. It appears that bandwidth limiting of the IF signal (input to the discriminator) will destroy, in general, the basis on which (in theory) the discriminator can produce the correct results. That is, the original assumption requires that the mean value of the input frequency in the total IF bandwidth equal the true carrier frequency. It is not certain that if the signal is passed through a narrower bandpass filter, even though it be symmetrical, that the mean value of the resulting signal will necessarily equal the true carrier frequency.

A clear example to indicate that the mean value is not preserved can be determined from Figure 1a. If a zonal bandpass filter is placed about the carrier to include only the -5 kHz and +5 kHz sidebands, then for this case the average frequency will equal that of the -5 kHz sideband. This follows because of the so-called capture effect, (Reference 3) wherein the average frequency of the signal is equal to the stronger component (in this case, the lower sideband). We must then conclude in general that the discriminator bandwidth, in order to give correct results, must encompass its entire RF signal spectrum; and therefore, there is no theoretical advantage gained by reducing the discriminator input bandwidth.

2.5 Use of Type 2 Information

To this point, we have looked to the goal of establishing a theoretical solution for the AM and PM cases by use of Type 1 information only, and found:

3. Downing, John J., "Modulations Systems and Noise", Prentice-Hall, Inc., Englewood Cliffs, N. J., 1964.

for AM: (a) Theoretically possible by determining center of the RF spectrum.

for PM: (a) Theoretically possible by use of perfect discriminator. However, operation is limited to carrier-to-noise ratio in order of -10 db in the discriminator input bandwidth.

(b) No theory is available which would lead to the determination of the carrier frequencies from the spectrum, or from an abbreviated portion of the spectrum.

Although, for the PM case, theory does not support a general solution for operation over the required range of signal-to-noise ratios, it appears that a feasible system may be possible if Type 2 information is made available. An obvious solution is the use of a computer to provide data reduction at the receiver, with appropriate checking against expected results to insure that a false lock has not occurred. This appears to be outside the scope of this effort. However, two other possibilities can be concluded:

1. If the RF spectrum is known to be completely symmetrical.

2. If the spectrum is not symmetrical, but is sufficiently well known to permit comparison with a priori spectral data.

With respect to the first possibility in which the symmetry of the input signal is guaranteed, a number of techniques can be used to automatically determine the center of the spectrum and, hence, determine the carrier frequency. What is not established here is whether or not the signals of interest are indeed sufficiently symmetrical. A study of this problem is indicated, but is beyond the scope of this effort. In this respect, however, it should be mentioned that the symmetry must initially be established at the spacecraft transmitter and preserved down through the IF amplifier of the ground receiver to the point where the spectrum is analyzed. Non-symmetry of the associated circuitry would no doubt affect the overall performance of such a system. This requirement for symmetrical bandpass characteristics in the receiver might well prove this system impractical.

The second possible system would be based on some sort of spectrum matching technique. Required here is a study of the spectrum of the various satellite transmitters to determine whether or not the carrier could be accurately identified. This, too, is beyond the scope of this effort, and no basis for predicting the performance has been established.

2.6 Spectrum Determination

For both of the possibilities suggested, a common equipment problem exists and can be broken down into two functions:

1. Fast and accurate determination of the signal spectrum.

2. Determination of the carrier frequency from the spectrum and implementing phase-lock.

A number of methods of establishing the spectra have been considered. These include the swept frequency approach, the comb filter, and the ubiquitous filter. The swept frequency analyzer is perhaps the simplest and cheapest method of obtaining the spectra, involving use of a single narrow bandpass filter and a swept frequency heterodyning circuit. If we assume an analysis bandwidth B equal to the phase-lock loop bandwidth and use a sweeping rate of B^2 Hz/second, then the time to analyze the entire 200 kHz spectrum is 200 kHz/B^2. Values for the various loop bandwidths are below:

Bandwidth	Analysis Time
3	20,000 seconds
10	2,000 seconds
30	200 seconds
100	20 seconds
300	2 seconds

It is at once seen that the required acquisition time of one second can be approached only for the 300-Hz loop bandwidth case. The analysis time can be speeded up by use of multiple (comb) filters; the time reduction factor being the number of filters employed. For example, the 30-Hz bandwidth would require a total of 200 filters to reduce the detection and acquisition time to approximately one second.

A third analyzer, the Federal Scientific Corporation Ubiquitous Spectrum Analyzer, has a performance roughly equivalent to a comb filter. It operates by digitizing and storing a 50 millisecond time sample of a 10 kHz band of information and then reads out this sample at a greatly increased rate to permit use of a wider bandwidth bandpass filter. In effect, the signal time base is contracted and, consequently, the spectrum is dilated. In order to analyze the required 200-kHz frequency band, twenty sequenced steps, each analyzing a 10-kHz band, would be required; the total time taking about two seconds. These heterodyning steps could be accomplished by a programmable frequency source such as a Hewlett-Packard frequency synthesizer.

2.7 Determination of Carrier from the Spectrum

Regardless of the scheme used to obtain the analysis, the carrier component must be determined. As has been stated, this can be done for AM by finding the center of symmetry of the spectrum. Several approaches to solving this problem exist using analog and/or digital techniques.

For the PM case, as has been mentioned, the center of symmetry may turn out in practice to provide a fairly accurate representation of the carrier frequency. However, in the event it does not, the problem of spectrum matching against known data could be considered. For this situation, a purely digital approach appears more attractive. Here, spectral data would be stored in the memory of a digital computer. The receiver IF signal would then be analyzed and inserted into the computer memory. Thereafter, the computer would be programmed to perform the required matching operation to determine the carrier. Once the carrier was found, it would be necessary to utilize the information in the receiver's local oscillator to achieve lock. In the event that the spectrum were unknown, the computer could perform other types of analysis such as determining the center of power or perhaps, for example, determine the average frequency of all spectra exceeding a given threshold (equivalent to the "punch thru" system spelled out in Electrac Proposal E.I. 391, dated 6/2/66). However, it must be emphasized that no theory has been found to assure the success of these methods.

2.8 Programmed VCO

A system has been brought to our attention by Mr. Victor R. Simas in which the VCO of the receiver is programmed to its correct lock frequency. The necessary information is derived from computer calculations of the Doppler shift on the transmitter carrier. The computer is presumably fed tracking data on each satellite and its transmitter frequency. It in turn supplies sample points of the predicted Doppler shift. This is done sufficiently advanced in time to permit this information to be used with a special analog programmer for the receiver VCO. This system, using Type 2 information, appears quite attractive compared to the other approaches suggested herein in that the information directly, rather than indirectly, concerns the carrier itself. However, as has been pointed out to us, the system does depend on predicting the frequency transmitted by the spacecraft transmitter which in practice could have considerable error due to drift. Whether or not this drift could be controlled or compensated for is not certain. If it can, this appears to be the most promising method to come to our attention for this problem, at least with respect to the use of Type 2 information.

III -- RECOMMENDED SYSTEM

Because a general solution of the problem of fully automatic acquisition of the carrier has not been found, consideration has been directed to the simpler problem of improving the present acquisition system. In this present system, the station operator searches the signal spectrum and decides which of one or more spectral lines represent the true carrier. This decision is based on his experience using the available aural and visual clues. In discussions with GSFC technical personnel, it was brought out that this procedure was too lengthy and requires continuous monitoring on the part of the station operator after lock-up. To provide an answer to this problem, a system has been devised here which provides the following features:

1. A system for rapidly and accurately establishing the significant spectral lines contained in the signal spectrum.

2. A system for displaying these lines to enable the operator to evaluate the selection problem.

3. Simple controls to initiate automatic acquisition of the phase-lock demodulator system to the selected spectral line.

4. A system which monitors the spectrum after lock-up and includes an alarm to warn the operator if more than one possible spectral line exists (if so, a false lock to a sideband or RFI from another satellite is possible).

The basic specifications recommended are:

Input Frequency Range:	140 kHz
Minimum Loop Bandwidth:	100 Hz
Lock Threshold:	-141 dbm*
Acquisition Time:	10 seconds (maximum)

3.1 Description of System

A block diagram of a system to accomplish this is shown in Figures 4 and 5. A bank of 14 spectral line detectors are provided, each of which covers a 10-kHz contiguous segment of the 140-kHz input bandwidth. Each detector contains a phase-lock loop tracking filter. All of these filters are continuously swept through their designated frequency band until such time as a signal is acquired. At the end of 10 seconds (the minimum time required to simultaneously sweep all 14 VCO's through their frequency range), as many as 14 spectral lines can be locked to based on one spectral component in each 10-kHz frequency band.

*Assumes 4 db receiver noise figure, 20 degrees rms loop noise

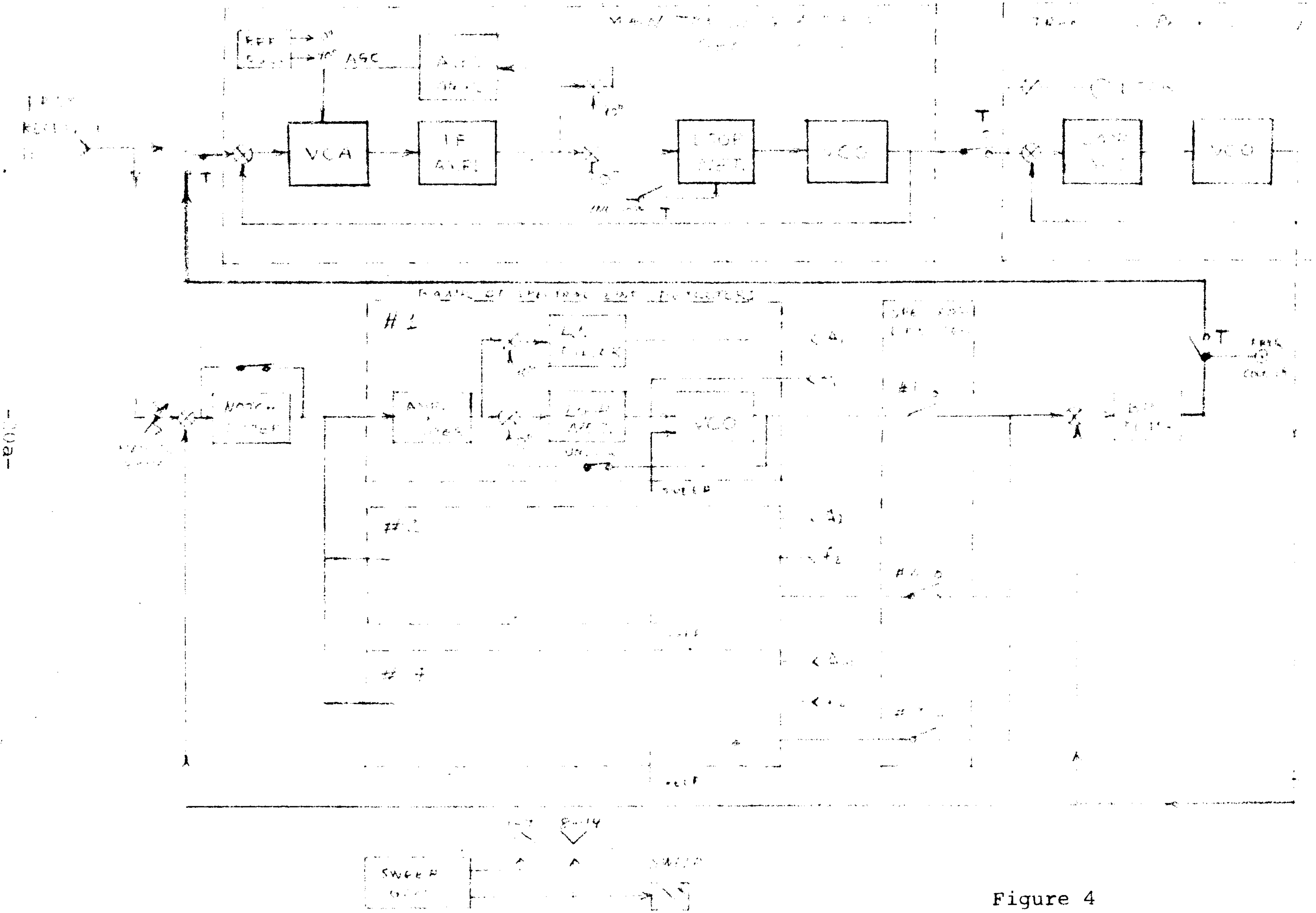


Figure 4

ACQUISITION AND MONITORING SYSTEM

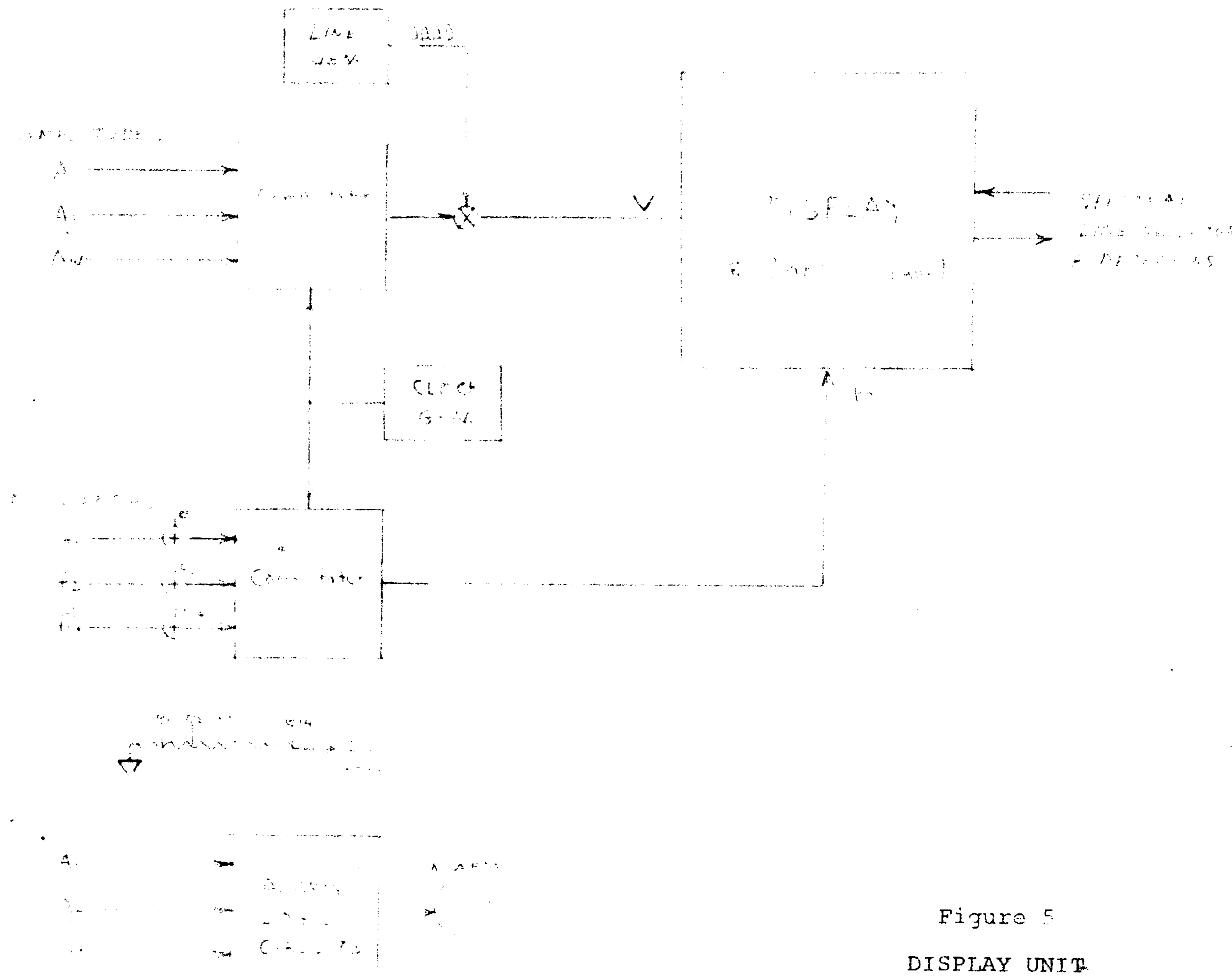


Figure 5

DISPLAY UNIT

The amplitude (A) and frequency (f) of each acquired spectral line is displayed on a large screen oscilloscope in the form of a bar-type graph (see Figure 6). The operator has available to him 14 switches; one for each tracking filter and its acquired spectral line. The operator selects one of these switches as the carrier. Subsequent depression of the transfer switch (T) transfers the VCO of the main tracking and data loop to track the selected spectral line. Detailed operation of the system is as follows:

1. Before the system mode switch is placed in the transfer (T) position, the VCO frequency of the main tracking and data loops is temporarily "memorized" by the transfer phase-lock loop.

2. The operator chooses a spectral line (for the carrier) by selection of the corresponding switch. The mode switch is then depressed to the verify position which perturbs the amplitude of the selected line presented on the display screen permitting the operator to verify he has selected the desired line.

3. The mode switch is then placed in the T position. The output of the bank of spectral line detectors consists of a single CW signal equal in frequency to the desired frequency component in the receiver IF.

4. The main tracking loop bandwidth is momentarily greatly increased and very rapid acquisition of the single CW spectral line from the detector bank output takes place. Using a 10-kHz loop bandwidth, the main tracking loop can be easily swept and locked in less than one second. During this period, the transfer phase-lock loop is not affected -- its basic purpose being to maintain the previously memorized VCO frequency for use in the spectral line detectors.

5. Return of the mode switch to its normal monitor position then accomplishes the following:

   a. The main tracking and data loop bandwidth is returned to its normal (100 Hz) bandwidth and it immediately (less than 0.1 second) acquires the selected frequency component in the receiver IF. This follows because the VCO of this loop is at precisely correct frequency to lock to the selected IF signal. Demodulation of the signal then takes place in the normal manner.

   b. The transfer loop re-acquires and then tracks the main tracking and data VCO for future memory action. This acquisition can take place very rapidly as no SNR or multiple choice problem exists.

c. The sweep-search cycle for all spectral line detectors again is repeated. Both the main tracking VCO and the transfer VCO track the desired input frequency and, hence, all frequencies presented on the display screen are indicated <u>relative</u> <u>to</u> the selected and tracked frequency. Again, a total of up to 10 seconds is required to completely search the 140-kHz band. (Note that the selected carrier can be tracked over ± 70 kHz range and, thus, the spectral line detectors -- because they operate relative to the carrier -- actually monitor a range of ± 140 kHz.)

6. If after lock-up the operator decides his choice was incorrect and wishes to select a different signal for the carrier, he merely presses the appropriate switch and repeats the stated procedure.

7. If the operator is satisfied with his selected frequency and it is the only one which exists, he then activates the alarm circuit which includes a notch filter to eliminate the desired signal to the detector bank input. Thereafter, the sweep search continues and, if a single line is later detected, the alarm system warns the operator of a possible false lock. If more than one signal can be locked to, he can

stand by and monitor the overall performance on the display unit. (Note: The notch filter has another important feature in that it will permit acquisition of signals having sidebands closely spaced to the carrier. By locking and notching out sidebands on a successive basis, the system can proceed from sideband to sideband until the true carrier is found.)

3.2 Summary of Performance of Spectral Line Detector Loops

Antenna Signal Level	Loop BW	SNR In 10 kHz	SNR, Loop*
-141 dbm	100 Hz	-11 db	+ 9 db
-135 dbm	155 Hz	- 5 db	+13 db
-130	270 Hz	0 db	+15.7 db
-120	300 Hz	+10 db	+25.2 db
-110	300 Hz	+20 db	+35.2 db
-100	300 Hz	+30 db	+45.2 db

System: Second-order; limiter IF

Sweep Speed: 1 kHz/second

Frequency Range: 10 kHz

3.3 Performance of Transfer Phase-Lock Loop

Acquisition Time: 1 Second maximum

Memory Performance, Input Removed = Drift <10 Hz/second

*20 Degrees rms noise jitter

3.4 Performance of Main Tracking Data Loop

Time to acquire spectral line detector band:	1 second maximum
Added time to acquire desired signal:	0.1 second
Sensitivity:	-141 dbm, 100 Hz bandwidth

3.5 Notch Filter

Rejection of center IF frequency:	≧ 50 db
3-db Bandwidth:	±25 Hz
1-db Bandwidth:	±50 Hz

3.6 Miscellaneous Design Features

a. Provide verify switch to momentarily eliminate a given line displayed on the display unit to insure the operator has selected the correct pushbutton.

b. Screen to be calibrated vertically 0-10 and the horizontal axis to be calibrated in kHz.

c. The horizontal scale of the display unit can be calibrated in 0 ±70 kHz. The beam is stepped in increments for each pulse of the clock generator.

d. The vertical scale of the display unit is calibrated 0-10 (db). Because of the requirement that at least 10 percent of the total power must reside in the true carrier, it follows that no other spectral line can be more than 90 percent (or by comparison 9.5 db).

e. If the commutator clock rate is set at 700 Hz, each vertical beam will be repeated 50 Hz/second, enough to preclude flicker effects.

f. The line generator can be a half-wave rectified 70-kHz sawtooth wave. This will provide approximately 100 vertical oscillations at each frequency point to insure an easily read bar-type indication.

g. A reset button can be added, if it is desired, to start all sweeps at the edge of each 10-kHz band. Further, it appears desirable to sweep loop 1-7 upward and loop 8-14 downward in frequency (always toward the center).

h. As an option, it may be desirable to add an additional #7 and #8 unit which would sweep in a direction away from the center to detect a possible second pair of signals which may exist close to the carrier.

i. Another option easy to accomplish is a digital counter read-out. In the verify position of the mode switch, a counter can be used to measure the absolute frequency difference between the selected carrier and the line of interest.

j. The manual gain control is to be used for very strong signals. Otherwise, the limiter will prevent relative spectral line amplitudes from being observed in the display unit (due to the fact that the limiter acts to produce a constant amplitude out of each detector for strong signals).

Figure 6 shows a possible panel layout of the recommended system. A number of variations are possible to include additional information to the operator before he makes his selection, including aural and visual outputs from the spectral line detectors similar to that used in the present phase-lock demodulators.

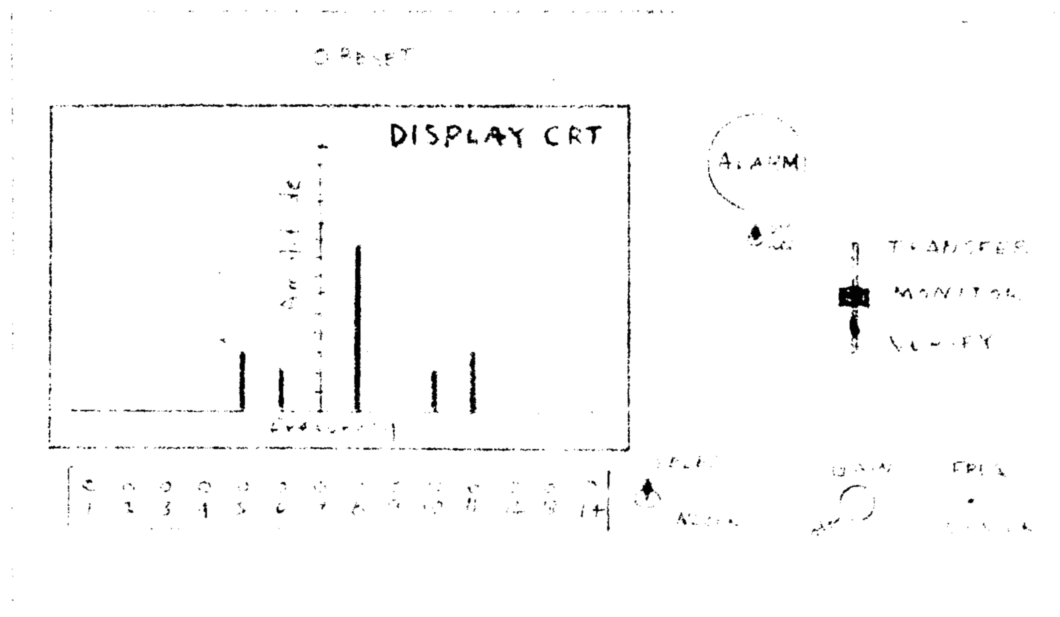


Figure 6

PANEL LAYOUT

IV -- CONCLUSIONS

Regarding the first problem of fully automatic acquisition:

1. The carrier frequency for AM can be found, both theoretically and practically by Type 1 information only, for all possible cases. The procedure follows directly from analysis of the signal spectrum.

2. The carrier frequency for PM can be found in theory, by Type 1 information only, for special cases in which the carrier-to-noise ratio exceeds approximately -10 db in the full spectrum bandwidth. The method requires an extraordinarily well designed discriminator. The use of spectral analysis has been ruled out from a theoretical standpoint, but may be helpful in practice.

3. For PM, in which the carrier-to-noise is below the stated -10 db value, Type 2 information is necessary.

4. Type 2 information in the form of a stored spectrum and a computer matching system has been suggested, but no information of feasibility is apparent without further study.

5. A system using Type 2 information in the form of predicted Doppler shift has been worked on at GSFC and appears most promising, but the feasibility may be limited because of transmitter drift.

As has been pointed out, the study was originally to be directed at the hardware required to achieve the desired stated goals. However, it became increasingly evident as the study progressed that a theoretical basis for positively identifying the carrier for PM (used to a significant extent on GSFC satellites) was just not available.

Regarding the second problem of semi-automatic acquisition: There is no question of the feasibility in solving the simplified problem to speed up the acquisition processes, wherein the station operator is required to make the final decision on whether or not the true carrier has been acquired.

V -- REFERENCES

1. Giacoletto, "Generalized Theory of Multitone Amplitude and Frequency Modulation", IRE Proceedings, (July, 1947), V. 35, p. 680.

2. Stumpers, F. L. H. M., "Theory of Frequency-Modulation Noise", IRE Proceedings, (September 1948), pp.1081-1092.

3. Downing, John J., "Modulations Systems and Noise", Prentice-Hall, Inc., Englewood Cliffs, N. J., 1964.

4. Goddard Space Flight Center Specification S-523-P-5, "Specifications for Automatic Acquisition Tracking Filter", Revised August 16, 1966.

APPENDIX I

DESIGN EVALUATION

OF THE

ANTI-SIDEBAND SYSTEM

FOR

DIVERSITY-LOCKED PHASE DEMODULATOR

CONTRACT NAS5-9864

12 AUGUST 1966

PREPARED BY

ELECTRAC, INC.

ANAHEIM, CALIFORNIA

FOR

GODDARD SPACE FLIGHT CENTER

GREENBELT, MARYLAND

SUMMARY

DESCRIPTION OF MODEL 315A
DIVERSITY LOCKED ANTI-SIDEBAND SYSTEM

The Diversity Lock Phase Demodulator includes features necessary for the minimization of sideband lock-on problems.

Modes of Operation: The anti-sideband lock circuit shall have two modes of operation, "Automatic" and "Normal". In the normal mode, the anti-sideband lock circuit shall serve only as an indicator. This circuit shall indicate whether the center frequency of the demodulator is above or below the center frequency of the desired signal (a signal with symmetrical spectral density is assumed).

In the automatic mode of operation, the anti-sideband lock circuit shall attempt to automate the acquisition function of the demodulator. In particular, when the acquisition mode is set below the center frequency of the desired signal, the following action shall take place:

1. Manual: Since this mode of operation is intended for threshold signals only, the sideband circuit shall have no effect.

2. Open: The sideband circuit shall have no effect.

3. Automatic: The anti-sideband lock circuit shall allow the operator to manually adjust the frequency control.

4. Sweep: The anti-sideband lock circuit shall allow the center frequency of the demodulator to sweep. But carrier lock-on must be 95 percent reliable. This mode of operation may be accomplished in any manner desired such as:

   a. Sweep-Lock-Unlock-Sweep; or

   b. a mode where discriminator sweep and auto-sweep are combined to maximum advantage.

Indicators: Three (3) indicator lights shall be provided to indicate anti-sideband system conditions as follows:

1. Automatic: To indicate when the sideband circuit is in the automatic mode.

2. Arrow Points Right: VCO above the center frequency of signal.

3. Arrow Points Left: VCO below the center frequency of signal.

4. <u>Both Arrows Indicate</u>: Anti-sideband circuit not activated due to low signal conditions.

5. <u>No Arrows Indicate</u>: VCO center frequency within 100 cycles of carrier.

DESIGN EVALUATION

OF THE

ANTI-SIDEBAND SYSTEM

FOR

DIVERSITY-LOCKED PHASE DEMODULATOR

INTRODUCTION

Reference should be made to Electrac Proposal E.I. 386 dated 28 April 1966 for development of an improved anti-sideband system for the Diversity-Locked Phase Demodulator. In this proposal, we discussed a basic system wherein two discriminators are used to center the VCO tuning to the approximate center of the signal spectrum prior to activation of existing phase-lock operation. A wideband discriminator of approximately 300 kHz is first used, and the result then refined by a 20 kHz narrow-band discriminator. Selection of circuit parameters to achieve stable operation of the AFC system consistent with reasonably low pull-in time for the desired accuracy is discussed.

NOMENCLATURE

f_i = Input Frequency (IF Carrier), Hz

f_o = Output Frequency, Hz

f_e = $f_i - f_o$ = Departure of VCO from Carrier Frequency, Hz

K_5 = Discriminator Sensitivity, Hz/volt

Y_3 = Transfer Function of Discriminator Low Pass Filter

K_4 = Common Channel VCO Sensitivity, radians/sec/volt

R_d = Feedback Resistor, Common Channel

C_b = Feedback Capacitor, Common Channel

See Reference 1

B_i = Discriminator Pre-Detection Bandwidth, Hz

C_i/N_i = Carrier to Noise Power Ratio in Discriminator Pre-Detection Bandwidth

1. Design Study Report For Diversity-Locked Phase Demodulator, 18 August 1965, Contract NAS5-9864, Electrac, Inc., Anaheim, California.

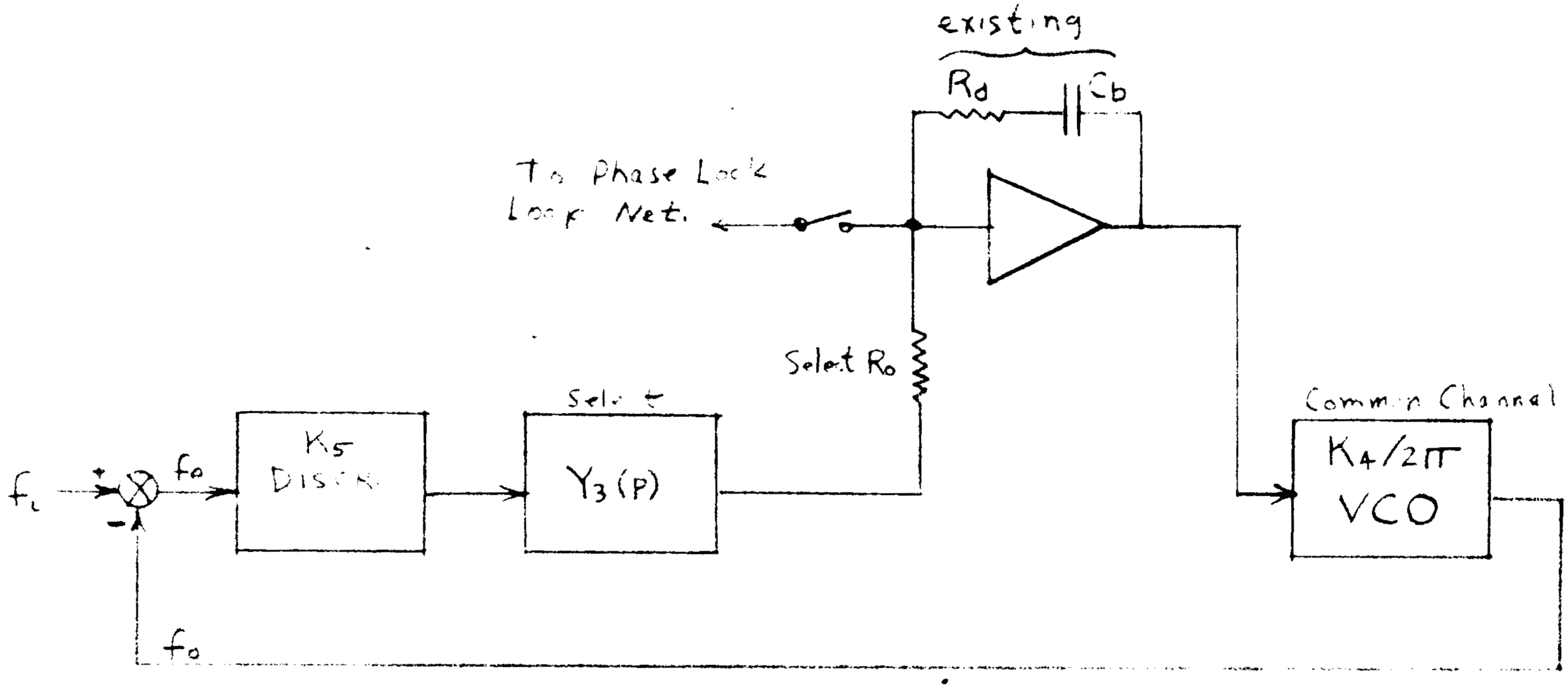


Figure 1

Basic AFC System

BASIC SYSTEM

The basic Anti-Sideband System utilizes a conventional AFC system as shown in Figure 1. The output of the discriminator contains a post-detection low-pass filter defined by $Y_3(p)$. This filter is designed to operate into the summing junction of the operational amplifier associated with the common channel VCO phase-lock tracking loop. This is done in order that the frequency of the VCO will be "memorized" when subsequent switching takes place from AFC to APC operation. The analysis which follows is based on an unmodulated carrier; no consideration has been given to a modulated signal.

STABILITY AND LOOP DYNAMICS

The appropriate applicable open loop equation can be described by inspection of Figure 1 as:

$$\frac{f_o}{f_e}(p) = K_5 \, Y_3 \, Y_4 \, \frac{K_4}{2\pi} = \frac{A(p)}{B(p)} \tag{1}$$

with $Y_4 = \frac{R_d}{R_o} \frac{p + \omega b}{p}$ where $\omega_b = 1/R_d C_b$ (1a)

$$\text{also,} \quad f_e(p) = f_i(p) - f_o(p) \tag{2}$$

From these equations, it is easy to show the closed loop relationships are:

$$\frac{f_o}{f_i}(p) = \frac{A}{A + B} \tag{3}$$

and
$$\frac{f_e}{f_i}(p) = \frac{B}{A + B} \tag{4}$$

It will later be shown that the simpliest practical filter following the discriminator must be a double pole network which we have chosen as follows:

$$Y_3(p) = \left(\frac{\omega_A}{p + \omega_A}\right)^2$$

Parameters previously established and constrained by readily available components are listed below for the wide and narrow bandwidth discriminator considered.

K_5 Discriminator: 10^{-5} V/Hz, Wide
10^{-3} V/Hz, Narrow

Filter Y_4:

Loop BW, Hz	R_d, Ohms	ω_b, Rad/Sec
10	155K	6.45
30	52K	19.5
100	15K	64.5
300	5K	195.

VCO: $K_4/2\Pi = 3 \times 10^4$ Hz/v

There remains the problem of choosing values ω_A and R_o to achieve a practical compromise of the following desired factors:

- Minimum settling time
- Minimum output frequency jitter (due to system noise)
- Dynamic stability
- Practical circuit parameter
- To minimize effect of beat frequency caused by beat of two differential-phase VCO's

NOISE

Before proceeding with the selection of ω_A and R_o, we determine first the noise jitter expected on f_o as follows: The noise power spectral density at the output of the discriminator can be shown to be

$$G_n(f) = \frac{f^2}{B_i F} \qquad \text{with } F = F(c_i/n_i) \tag{5}$$

where c_i/n_i = carrier to noise power ratio in IF and F is found from Figure 2

and $$\overline{f}_o^{\,2} = \int_0^{B_i} G_n(f) \left| \frac{f_o}{f_i}(jf) \right|^2 df$$ where $\overline{f}_o^{\,2}$ = mean square jitter

Our choice of Y_3 is such that it effectively establishes an upper bound on the noise output and, therefore, for simplicity we can write

$$\overline{f_o}^2 \leq \int_o^{B_i} G_n(f)\; Y_3 Y_3^* \, df \tag{6}$$

Substituting

$$\overline{f_o}^2 \leq \frac{1}{B_i F} \int_o^{B_i} f^2 \left(\frac{f_A^2}{f^2 + f_A^2} \right)^2 df$$

$$\overline{f_o}^2 \leq \frac{f_A^3}{B_i F} \frac{\pi}{4} \qquad \text{for } B_i >> f_A$$

and the rms frequency jitter $\tilde{f}_o$ is

$$\tilde{f}_o = f_A^{3/2} \left(\frac{\pi}{4 B_i F} \right)^{½} \tag{7}$$

For the wideband discriminator, $B_i \pm 3 \times 10^5$ Hz. The system AGC operates on the carrier down to approximately zero db carrier to noise ratio in the IF amplifier bandwidth. At this point, the conditions in the receiver system will be in the following Table 1.

TABLE I

IF BW	ANTENNA SIGNAL	IF CARRIER	IF NOISE	IF NOISE DENSITY	NOISE INPUT 300 kHz DISC	C_i/N_i	$F(C_i/N_i)$*	$\tilde{f}_o$,rms $\leq$
3 MHz	-106 dbm	-13 dbm	-13 dbm	-78 dbm/Hz	-23 dbm	+10 db	+9.2 db	.002 Hz
1 MHz	-111	-13	-13	-73	-18	+ 5	+1.6	.005
300 kHz	-116	-13	-13	-68	-13	0	-10.3	.02
100 kHz	-121	-13	-13	-63	-13	0	-10.3	.035
30 kHz	-126	-13	-13	-58	-13	0	-10.3	.06
10 kHz	-131	-13	-13	-53	-13	0	-10.3	.1

*Middleton, D., An Intro. to Stat. Comm. Theory, Chapter 15.

To gain some insight into the expected noise, the last column in the above table is the calculated rms frequency jitter for an arbitrary selected value of $f_A = 15/2\Pi = 2.4$ Hz. It is clear that the resulting VCO jitter for SNR = 0 in the IF bandwidth is negligible, and the final accuracy should be only a matter of zero drift in the circuits.

Selected values,

$\omega_A = 15$ radians/sec

$R_o = 51$K ohms, wideband

5.1M ohms, narrow-band

$K_5 = 10^{-5}$ v/Hz wideband

$K_5 = 10^{-3}$ v/Hz narrow-band

$\frac{K_4}{2\Pi} = 3 \times 10^4$ Hz/v

From equation 1, 2, and 3

$$\frac{f_o}{f_i} = \frac{K\omega_A^2 \ (p + \omega b)}{p^3 + 2\ \omega_A p^2 + (K + 1)\omega_A^2 p + K\omega_A^2 \omega_b}$$

where $K = \frac{K_4 K_5 R_d}{2\Pi R_o}$ (identical for both wide and narrow-bandwidth)

$$\frac{f_e}{f_i} = \frac{p(p + \omega_A)^2}{p^3 + 2\omega_A p^2 + (K + 1)\omega_A^2 p + K\omega_A^2 \omega_b}$$

Loop BW	ω_b	R_d	K
10 Hz	6.45	155×10^3	.91
300 Hz	195	5×10^3	.0294

For the 10 Hz loop,

$$\frac{f_o}{f_i} = \frac{205(p + 6.45)}{(p + 4.1)(p^2 + 2\xi\omega_n p + \omega_n^2)} \qquad \omega_n = 18, \quad \xi = .72$$

and

$$\frac{f_e}{f_i} = \frac{p(p + 15)^2}{(p + 4.1)(p^2 + 2\xi\omega_n p + \omega_n^2}$$

$$\simeq \frac{p}{p + 4}$$

and for the 300 Hz loop,

$$\frac{f_o}{f_i} = \frac{6.85(p + 195)}{(p + 22.3)(p^2 + 2\xi\omega_n p + \omega_n^2)} \qquad \omega_n = 7.7 \text{ rad}, \quad \xi = 0.5$$

and

$$\frac{f_o}{f_i} = \frac{p(p + 15)^2}{(p + 22.3)(p^2 + 2\xi\omega_n p + \omega_n^2)}$$

or $\approx \frac{p}{p + 8}$

Using the approximate error transfer function, we note the frequency vs. time transient for a step value of F_i as follows:

10 Hz Loop:

$$f_e(f) = F_i e^{-4t}; \text{ and } \dot{f}_e = -4\ f_e(t)$$

and for the 300 Hz Loop:

$$f_e(t) = F_i e^{-8t}; \ \dot{f}_e = -8\ f_e(t)$$

We now compute the time for the error to settle to the 5 kHz and 50 Hz values for the wide and narrow bandwidths, respectively, and the frequency rate:

SETTLING TIME

Loop Bandwidth	Wideband 100 to 5 kHz	Narrow-Band 5 to 50 Hz	T Total Time	$\dot{f}_e$ (T)
10 Hz	0.8 sec	1.2 sec	2 sec	-200 Hz/sec
300 Hz	0.4 sec	.6 sec	1 sec	-400 Hz/sec

It appears a maximum of two seconds will be required to achieve the desired results. Also, assuming the relay logic circuitry requires a maximum of 5 milliseconds, we see that the overshoot will not exceed 2 Hz of the transfer from AFC to phase lock operation.

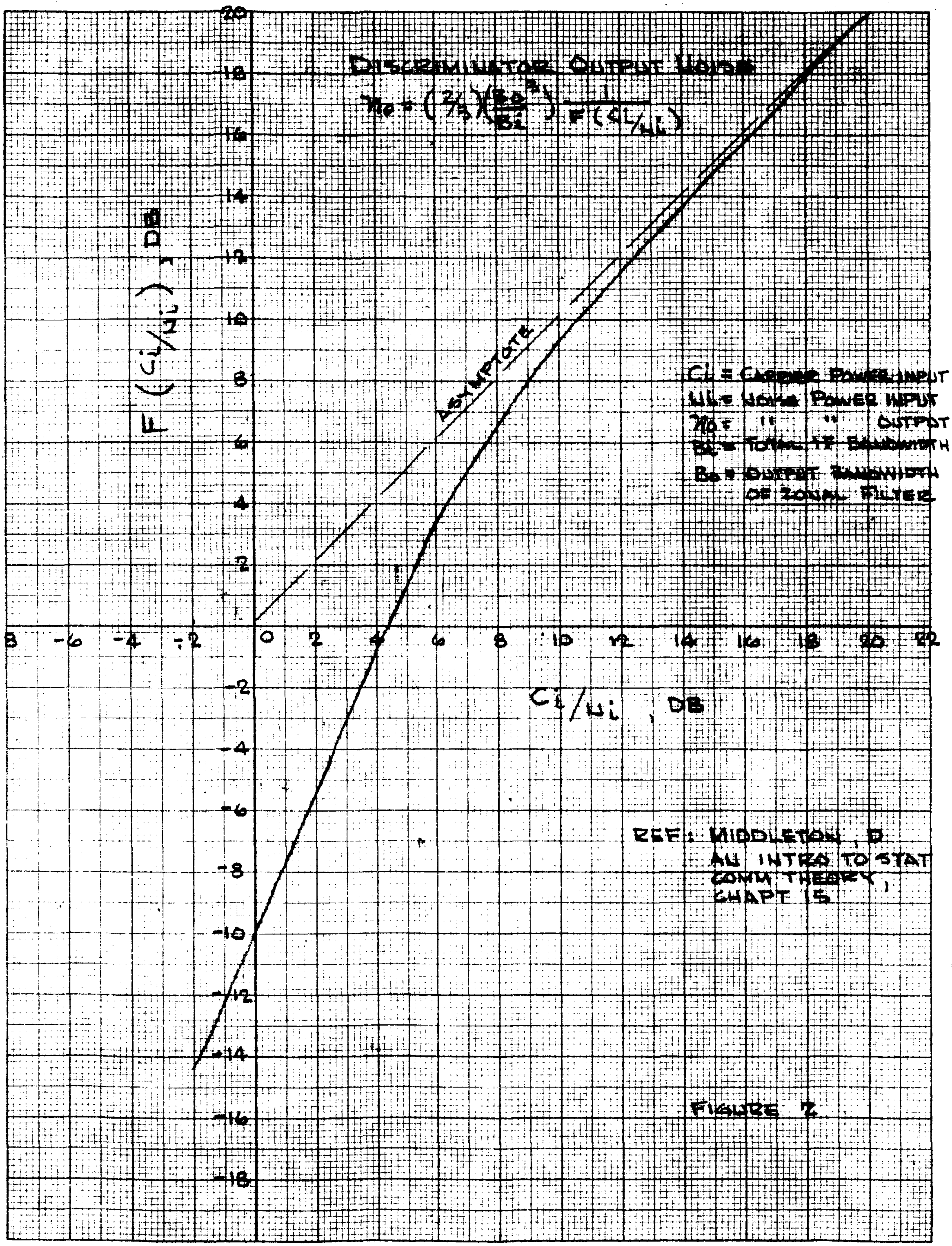
RELIABILITY

Signal strength is reliably indicated by the AGC voltage for receiver input signal-to-noise ratios greater than unity measured in the IF bandwidth. For SNR's less than unity, the AGC voltage is constant and therefore does not indicate the signal level. It is clear from Table I that all receiver input signal levels > 106 dbm can be reliably measured. If this level point is used to automatically calibrate the system, AFC would be enabled for SNR exceeding that listed in Table II below:

IF BW	SNR in IF BW	C_i/N_i 300 kHz Disc.	C_i/N_i 20 kHz Disc.
3 MHz	0 db	10	+22
1 MHz	5	10	+22
300 kHz	10	10	+22
100 kHz	15	15	+22
30 kHz	20	20	+22
10 kHz	25	25	25

TABLE II

From our previous calculations on noise jitter, it appears the AFC is capable of operation down to at least zero db in all IF bandwidths. A method for achieving this is to scale the AGC voltage coincident with the setting of the receiver IF bandwidth switch. This would be accomplished indirectly by utilizing the demod-bandwidth switches instead, with the understanding that the station operators would be required to use corresponding settings on with the receiver and demodulator switches. The feasibility of utilizing this approach requires further discussions with GSFC personnel.



DISCRIMINATOR OUTPUT NOISE
$N_o = \left(\frac{2}{3}\right)\left(\frac{B_o^3}{B_i}\right)\frac{1}{F(C_i/N_i)}$
ASYMPTOTE
$F\left(C_i/N_i\right)$, DB
C_i/N_i, DB
C_i = CARRIER POWER INPUT
N_i = NOISE POWER INPUT
N_o = " " OUTPUT
B_i = TOTAL IF BANDWIDTH
B_o = OUTPUT BANDWIDTH OF ZONAL FILTER
REF: MIDDLETON, D. AN INTRO TO STAT COMM THEORY, CHAPT 15
-8 -6 -4 -2 0 2 4 6 8 10 12 14 16 18 20 22
20 18 16 14 12 10 8 6 4 2 -2 -4 -6 -8 -10 -12 -14 -16 -18

FIGURE 7